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Hybrid Schemes for Adaptive Control Strategies

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1. Introduction

The purpose of this chapter is to redesign the standard adaptive control schemes by using hybrid structure composed by Model Reference Adaptive Control (MRAC) or Adaptive Pole Placement Control (APPC) strategies, associated to Variable Structure (VS) schemes for achieving non-standard robust adaptive control strategies. The both control strategies is now on named VS-MRAC and VS-APPC. We start with the theoretical base of standard control strategies APPC and MRAC, discussing their structures, as how their parameters are identified by adaptive observers and their robustness properties for guaranteeing their stability. After that, we introduce the sliding mode control (variable structure) in each control scheme for simplifying their design procedure. These design procedure are based on stability analysis of each hybrid robust control scheme. With the definition of both hybrid control strategies, it is analyzed their behavior when controlling system plants with unmodeled disturbances and parameter variation. It is established how the adaptive laws compensates these unmodeled dynamics. Furthermore, by using simple systems examples it is realized a comparison study between the hybrid structures VS-APPC and VSMRAC and the standard schemes APPC and MRAC. As the hybrid structures use switching laws due to the sliding mode scheme, the effect of chattering is analyzed on the implementation and consequently effects on the digital control hardware where sampling times are limiting factor. For reducing these drawbacks it is also discussed possibilities which kind of modifications can employ. Finally, some practical considerations are discussed on an implementation on motor drive systems.

2. Variable Structure Model Reference Adaptive Controller (VS-MRAC)

The *VS-MRAC* was originally proposed in (Hsu *et al.*, 1989) and extensively discussed in (Hsu *et al.*, 1994). The main features of this control scheme are the robustness of parameters uncertainties and unmodeled disturbances, as well as good transitory response.
Consider the following first order plant

$$W(s) = \frac{y}{u} = \frac{b_p}{s + a_p} \ , \tag{1}$$

where b_p and a_p are unknown or known with limited uncertainties. Admitting a reference model given by

$$M(s) = \frac{y_m}{r} = \frac{b_m}{s + a_m}, \tag{2}$$

in which $k_m > 0$ and $a_m > 0$, the following output error variable can be defined as

$$e_0 = y - y_m \,. \tag{3}$$

The control objective is to force $y(t)$ to asymptotically track the reference output signal, $y_m(t)$, by regulating e_0 to be zero, while keeping all the closed-loop signals uniformly bounded. The control law used for accomplished this is

$$u = \theta_1 y + \theta_2 r \,, \tag{4}$$

which is the same as used in traditional model reference adaptive control. However, instead of the integral adaptive laws for the controller parameters, switching laws are proposed in order to improve the system transient performance and its robustness.

If b_p and a_p are known, the ideal controller parameters (θ_1^* and θ_2^*) can be founded using the following condition

$$\frac{y}{r} = \frac{y_m}{r} \,, \tag{5}$$

which means that our control objective is achieved, i.e., the closed-loop system behaves like the open-loop reference model. Consequently, the control law equation can be rewritten as

$$u = \theta_1^* y + \theta_2^* r \,. \tag{6}$$

Analyzing (1) and (2) in the time domain, we get

$$\dot{y} = -a_p y + k_p u \,, \tag{7}$$

$$\dot{y}_m = -a_m y_m + k_m r \,. \tag{8}$$

Adding and subtracting terms related to the ideal control parameters in (4),

$$u = \theta_1 y + \theta_2 r - \theta_1^* y - \theta_2^* r + \theta_1^* y + \theta_2^* r \,, \tag{9}$$

and then grouping some terms

$$u = (\theta_1 - \theta_1^*)y + (\theta_2 - \theta_2^*)r + \theta_1^* y + \theta_2^* r\,, \tag{10}$$

we have

$$u = \tilde{\theta}_1 y + \tilde{\theta}_2 r + \theta_1^* y + \theta_2^* r\,, \tag{11}$$

in which terms $\tilde{\theta}_1$ and $\tilde{\theta}_2$ are deviations of ideal controller parameters θ_1 and θ_2. Substituting the resulting equation (11) in (7),

$$\dot{y} = -a_p y + b_p(\tilde{\theta}_1 y + \tilde{\theta}_2 r + \theta_1^* y + \theta_2^* r)\,, \tag{12}$$

we can rewrite this equation as

$$\dot{y} = -a_p y + b_p\theta_1^* y + b_p\theta_2^* r + b_p(\tilde{\theta}_1 y + \tilde{\theta}_2 r)\,, \tag{13}$$

which results in

$$\dot{y} = -(a_p - b_p\theta_1^*)y + b_p\theta_2^* r + b_p(\tilde{\theta}_1 y + \tilde{\theta}_2 r)\,. \tag{14}$$

From (6), the model input r can be defined as

$$r = \frac{u - \theta_1^* y}{\theta_2^*}\,. \tag{15}$$

Therefore, using (11) and (15) in (8), we get

$$\dot{y}_m = -a_m y + b_m r + \frac{b_m}{\theta_2^*}(\tilde{\theta}_1 y + \tilde{\theta}_2 r)\,. \tag{16}$$

Finally, comparing (14) and (16) due to the condition (5), we have the desired controller parameters

$$\theta_1^* = \frac{a_p - a_m}{b_p}\,, \tag{17}$$

$$\theta_2^* = \frac{b_m}{b_p}. \tag{18}$$

The above desired controller parameters assure that plant output converges to its reference model, because b_p and a_p are known. This design criteria is named as *The Matching Conditions*.
However, our interests are concerned with unknown plant parameters or with known plant parameters with uncertainties, which require the use of adaptive laws for adjusting controller parameters. Derivating the output error equation given in (3),

$$\dot{e}_0 = \dot{y} - \dot{y}_m \tag{19}$$

and using the condition (5), with equations (8), (16) and (19), we get

$$\dot{e}_0 = -a_m y + b_m r + \frac{b_m}{\theta_2^*}(\tilde{\theta}_1 y + \tilde{\theta}_2 r) - (-a_m y_m + b_m r), \tag{20}$$

which can be rearranged as

$$\dot{e}_0 = -a_m(y - y_m) + \frac{b_m}{\theta_2^*}(\tilde{\theta}_1 y + \tilde{\theta}_2 r). \tag{21}$$

Thus,

$$\dot{e}_0 = -a_m e_0 + \frac{b_m}{\theta_2^*}(\tilde{\theta}_1 y + \tilde{\theta}_2 r). \tag{22}$$

Now, consider the Lyapunov function candidate given by

$$V(e_0) = \frac{1}{2}e_0^2 > 0, \tag{23}$$

and its respective first time derivative

$$\dot{V}(e_0) = e_0\dot{e}_0. \tag{24}$$

By substituting (22) in (24), we obtain the following equation

$$\dot{V}(e_0) = \left[-a_m e_0 + \frac{b_m}{\theta_2^*} (\tilde{\theta}_1 y + \tilde{\theta}_2 r) \right] e_0 \ , \tag{25}$$

that can be rewritten as

$$\dot{V}(e_0) = -a_m e_0^2 + \frac{b_m}{\theta_2^*} \Big[\Big(\theta_1 - \theta_1^* \Big) e_0 y + \Big(\theta_2 - \theta_2^* \Big) e_0 r \Big]. \tag{26}$$

Using the switching laws,

$$\theta_1 = -\overline{\theta}_1 sgn(e_0 y) \, , \tag{27}$$

$$\theta_2 = -\overline{\theta}_2 sgn(e_0 r) \, , \tag{28}$$

we obtain,

$$\dot{V}(e_0) = -a_m e_0^2 - \frac{b_m}{\theta_2^*} \Big[\Big(\overline{\theta}_1 \mid e_0 y \mid + \theta_1^* e_0 y \Big) + \Big(\theta_2 \mid e_0 r \mid + \theta_2^* e_0 r \Big) \Big]. \tag{29}$$

If the conditions $\overline{\theta}_1 >\mid \theta_1^* \mid$ and $\overline{\theta}_2 >\mid \theta_2^* \mid$ are satisfied, the terms with indefinite signals in (29) are dominated, and then

$$\dot{V}(e_0) \leq -a_m e_0^2 < 0 \tag{30}$$

which guarantees that $e_0 = 0$ is a globally asymptotically stable (GAS) equilibrium point, because (30) is a negative definite function.

3. Variable Structure Adaptive Pole Placement Control (VS-APPC)

As the VS-MRAC, the VS-APPC is the hybrid control structure obtained from the association of Pole Placement Control (PPC) together with Variable Structure (VS). Therefore, the theoretical development of this section starts from PPC control scheme and then we introduce the VS concepts for achieving the proposed VS-APPC.
Considering the single input/single output (SISO) LTI plant

$$y = G(s) u \, , \tag{31}$$

in which

$$G(s) = \frac{Z(s)}{R(s)} = \frac{b_{n-1}s^{n-1} + ... + b_1 s + b_0}{s^n + a_{n-1}s^{n-1} + ... + a_1 s + a_0} \ , \tag{32}$$

there are, as plant parameters, $2n$ elements, which are the coefficients of the numerator and denominator of transfer function $G(s)$. We can define the vector θ^* as

$$\theta^* = \begin{bmatrix} b_{n-1} & . & . & . & b_1 & b_0 & a_{n-1} & . & . & . & a_1 & a_0 \end{bmatrix}^T . \tag{33}$$

From this, the following constraints must be observed:

S1. $R(s)$ is a monic polynomial whose degree n is known.

S2. $Z(s)$, $R(s)$ are coprime and degree$(Z) < n$.

Assumptions (S1) and (S2) allow $Z(s)$, $R(s)$ to be non-Hurwitz in contrast to the MRC (Model Reference Control) case, where $Z(s)$ is required to be Hurwitz.

We can also extend the PPC scheme for including the tracking objective, where output y is required to follow a certain class of reference signals r, by using the internal model principle (Ioannou & Sun, 1996). The uniformly bounded reference signal is assumed to satisfy

$$Q_m(s)r = 0, \tag{34}$$

where $Q_m(s)$, the internal model of r, is a known monic polynomial of degree q with non-repeated roots on the $j\omega$-axis and satisfies

S3. $Q_m(s)$, $Z(s)$ are coprime.

Considering the control law given by

$$Q_m(s)L(s)u = -P(s)y + M(s)r \ , \tag{35}$$

where $P(s)$, $M(s)$, $L(s)$ are polynomials (with $L(s)$ monic) of degree $q+n-1$, $q+n-1$ e $n-1$, respectively, and $Q_m(s)$ satisfies (34) and assumption (S3).

Applying (35) to the plant (31), we obtain the closed-loop plant equation

$$y = \frac{Z(s)M(s)}{Q_m(s)L(s)R(s) + P(s)Z(s)} r \ , \tag{36}$$

whose characteristic equation is

$$Q_m(s)L(s)R(s) + P(s)Z(s) = 0, \tag{37}$$

and has order $2n + q - 1$. The objective now is chosen $P(s)$, $L(s)$ such that

$$Q_m(s)L(s)R(s) + P(s)Z(s) = A^*(s) \tag{38}$$

is satisfied for a given monic Hurwitz polynomial $A^*(s)$ of degree $2n + q - 1$. Because of assumptions S2 e S3 which guarantee that $Q_m(s)$, $R(s)$, $Z(s)$ are coprime, there is a solution so that $L(s)$ and $P(s)$ satisfy (38) and this solution is unique (Ioannou & Sun, 1996).
Using (38), the closed-loop is described by

$$y = \frac{Z(s)M(s)}{A^*(s)} r. \tag{39}$$

Similarly, from the plant (31) and the control law (35) and (38), we obtain

$$u = \frac{R(s)M(s)}{A^*(s)} r. \tag{40}$$

Because r is uniformly bounded and $\frac{Z(s)M(s)}{A^*(s)}$, $\frac{R(s)M(s)}{A^*(s)}$ are proper with stable poles, y and u remain bounded whenever $t \to \infty$ for any polynomial $M(s)$ of degree $n + q - 1$ (Ioannou & Sun, 1996). Therefore, the pole placement objective is achieved by the control law (35) without having any additional restrictions in $M(s)$ and $Q_m(s)$. When $r = 0$, (39) and (40) imply that y and u converge to zero exponentially fast. On the other hand, when $r \neq 0$, the tracking error $e = y - r$ is given by

$$e = \frac{Z(s)M(s) - A^*(s)}{A^*(s)} r = \frac{Z(s)}{A^*(s)}[M(s) - P(s)]r - \frac{L(s)R(s)}{A^*(s)} Q_m(s) r. \tag{41}$$

In order to obtain zero tracking error, the equation above suggests the choice of $M(s) = P(s)$ to cancel its first term, while the second term can be canceled by using (34). Therefore, the pole placement and tracking objective are achieved by using the control law

$$Q_m(s)L(s)u = -P(s)(y - r), \tag{42}$$

which is implemented as shown in Fig. 1 using $n+q-1$ integrators for the controller realization. An alternative realization of (42) is obtained by rewriting it as

$$u = \frac{\Lambda - LQ_m}{\Lambda} u - \frac{P}{\Lambda}(y-r)\,, \tag{43}$$

where Λ is any monic Hurwitz polynomial of degree $n+q-1$.

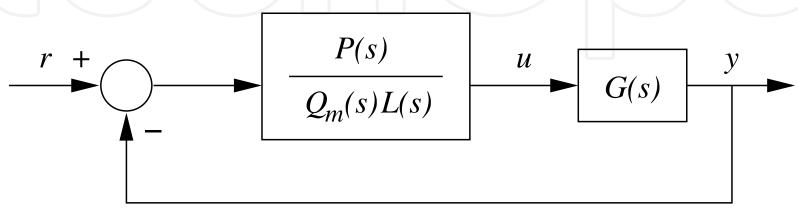


Fig. 1. Block diagram of pole placement control.

The PPC design supposes that the plant parameters are known, what not always is true or possible. Therefore, integral adaptive laws can be proposed for estimating these parameters and then used with PPC schemes. This new strategy is called Adaptive Pole Placement Controller (APPC), where the c*ertainty equivalence principle* guarantees that the output plant tracks the reference signal r, if the estimates converge to the desired values. In this section, instead of these traditional adaptive laws, switching laws will be used for the the first order plant case, according to (Silva *et al.*, 2004).
Consider the plant,

$$y = \frac{b}{s+a} u\,, \tag{44}$$

and its respective time domain equation,

$$\dot{y} = -ay + bu\,, \tag{45}$$

where the parameters a and b are unknown or known with uncertainties. Let be a_m a positive constant, we may write (45) by adding and subtracting the term $a_m y$,

$$\dot{y} = -a_m y + (a_m - a)y + bu\,. \tag{46}$$

A model for the plant may be written as

$$\dot{\hat{y}} = -a_m \hat{y} + (a_m - \hat{a})y + \hat{b}u\,, \tag{47}$$

where $\hat{a}$ and $\hat{b}$ are estimates for a and b, respectively (Ioannou & Sun, 1996).
We define the estimation error e_0 as

$$e_0 = y - \hat{y}\,, \tag{48}$$

and with (46) and (47), we get

$$\dot{e}_0 = -a_m e_0 + \tilde{a}y - \tilde{b}u\,, \tag{49}$$

where

$$\tilde{a} = \hat{a} - a\ , \tag{50}$$
$$\tilde{b} = \hat{b} - b\ . \tag{51}$$

Choosing the following Lyapunov function candidate,

$$V(e_0) = \frac{1}{2}{e_0}^2 > 0\,, \tag{52}$$

we have

$$\dot{V}(e_0) = e_0\dot{e}_0\ , \tag{53}$$

which can be rewritten using (49),

$$\dot{V}(e_0) = -a_m {e_0}^2 + \tilde{a}e_0 y - \tilde{b}e_0 u\,. \tag{54}$$

Expanding the above equation with (50) and (51),

$$\dot{V}(e_0) = -a_m {e_0}^2 + (\hat{a} - a)e_0 y - (\hat{b} - b)e_0 u\,, \tag{55}$$

and then using the switching laws,

$$\hat{a} = -\overline{a}\,\mathrm{sgn}(e_0 y)\,, \tag{56}$$
$$\hat{b} = \overline{b}\,\mathrm{sgn}(e_0 u)\,, \tag{57}$$

we get,

$$\dot{V}(e_0) = -a_m e_0^{\,2} - (\bar{a}\left|e_0 y\right| + a e_0 y) - (\bar{b}\left|e_0 u\right| - b e_0 u)\,. \tag{58}$$

Finally, if the conditions $\bar{a} > \left|a\right|$ and $\bar{b} > \left|b\right|$ are satisfied,

$$\dot{V}(e_0) \le -a_m e_0^{\,2} < 0\,, \tag{59}$$

which guarantees that $e_0 = 0$ is a globally asymptotic stable (GAS) equilibrium point. Moreover, if we follow a similar procedure described in (Hsu & Costa, 1989), we can prove that $e_0 = 0$ reaches the sliding surface in a finite time t_f ($e_0 = 0$, $\forall t > t_f$).

4. Application on a Current Control Loop of an Induction Machine

To evaluate the performance of both proposed hybrid adaptive schemes, we use an induction machine voltage x current model as an experimental plant. The voltage equations of the induction machine on arbitrary reference frame can be presented by the following equations:

$$v_{sd}^g = \left(r_s + \frac{l_s - \sigma l_s}{\tau_r}\right) i_{sd}^g + \sigma l_s \frac{di_{sd}^g}{dt} - \omega_g \sigma l_s i_{sq}^g - \left(\frac{l_s - \sigma l_s}{l_m}\right)\left(\omega_r \phi_{rq}^g + \frac{\phi_{rd}^g}{\tau_r}\right), \tag{60}$$

$$v_{sq}^g = \left(r_s + \frac{l_s - \sigma l_s}{\tau_r}\right) i_{sq}^g + \sigma l_s \frac{di_{sq}^g}{dt} + \omega_g \sigma l_s i_{sd}^g + \left(\frac{l_s - \sigma l_s}{l_m}\right)\left(\omega_r \phi_{rd}^g - \frac{\phi_{rq}^g}{\tau_r}\right), \tag{61}$$

where v_{sd}^g, v_{sq}^g, i_{sd}^g and i_{sq}^g are *dq* axis stator voltages and currents in a generic reference frame, respectively; r_s, l_s and l_m are the stator resistance, stator inductance and mutual inductance, respectively; ω_g and ω_r are the angular frequencies of the *dq* generic reference frame and rotor reference frame, respectively; $\sigma = 1 - l_m^2 / l_s l_r$ and $\tau_r = l_r / r_r$ are the leakage factor and rotor time constant, respectively.

The above model can be simplified by choosing the stator reference frame ($\omega_g = 0$). Therefore, equations (60) and (61) can be rewritten as

$$v_{sd}^s = r_{sr} i_{sd}^s + \sigma l_s \frac{di_{sd}^s}{dt} + e_{sd}^s\,, \tag{62}$$

$$v_{sq}^{s} = r_{sr} i_{sq}^{s} + \sigma l_{s} \frac{di_{sq}^{s}}{dt} + e_{sq}^{s} \;, \tag{63}$$

where s is the superscript related to the stator reference frame, $r_{sr} = r_{s} + (l_{s} - \sigma l_{s}) / \tau_{r}$, e_{sd}^{s} and e_{sq}^{s} are *fcems* of the *dq* machine phases given by

$$e_{sd}^{s} = -\left(\omega_{r}\phi_{rq}^{s} + \frac{\phi_{rd}^{s}}{\tau_{r}}\right)\frac{(l_{s} - \sigma l_{s})}{l_{m}}, \tag{64}$$

and

$$e_{sq}^{s} = \left(\omega_{r}\phi_{rd}^{s} - \frac{\phi_{rq}^{s}}{\tau_{r}}\right)\frac{(l_{s} - \sigma l_{s})}{l_{m}}, \tag{65}$$

The current x voltage transfer function of the induction machine can be obtained from (62) and (63) as

$$\frac{I_{sd}^{s}(s)}{V_{sd}^{s'}(s)} = \frac{I_{sd}^{s}(s)}{V_{sd}^{s'}(s)} = \frac{1 / r_{sr}}{s\tau_{s} + 1}, \tag{66}$$

where $\tau_{s} = \sigma l_{s} / r_{sr}$, $V_{sd}^{s'}(s) = V_{sd}^{s}(s) - E_{sd}^{s}(s)$ and $V_{sq}^{s'}(s) = V_{sq}^{s}(s) - E_{sq}^{s}(s)$. The *fcems* $E_{sd}^{s}(s)$ and $E_{sq}^{s}(s)$ are considered unmodeled disturbances to be compensated by the control scheme.

Analyzing the current x voltage transfer functions of a standard machine, we can observe that the time constant τ_{s} has parameters which vary with the dynamic behavior of machine. Moreover, this plant has also unmodeled disturbances. This justifies the use of this control plant for evaluating the performance of proposed control schemes.

5. Control System

Fig. 2 presents the block diagram of a standard vector control strategy, in which the proposed control schemes are employed for induction motor drive. Block *RFO* realizes the vector rotor field oriented control strategy. It generates the stator reference currents i_{sd}^{s*} and i_{sq}^{s*}, angular stator frequency ω_{o}^{*} of stator reference currents from desired reference torque T_{e}^{*}, and reference rotor flux ϕ_{r}^{*}, respectively. Blocks *VS-ACS* implement the proposed robust adaptive current control schemes that could be the *VS-MRAC* strategy or the *VS-*

APPC strategy. Both current controllers are implemented on the stator reference frame. Block $dq^s/123$ transforms the variables from dq^s stationary reference frame into 123 stator reference frame.
Generically, the current-voltage transfer function given by equation (66) can be rewritten as

$$W_{isdq}^s(s) = \frac{I_{sd}^s(s)}{V'^s_{sd}(s)} = \frac{I_{sq}^s(s)}{V'^s_{sq}(s)} = \frac{b_s}{s + a_s}, \tag{67}$$

in which $b_s = 1/\sigma l_s$ and $a_s = 1/\tau_s$. In this model, the *fcems* e_{sd}^s and e_{sq}^s are considered unmodeled disturbances to be compensated by current controllers. The parameters a_s and b_s are known with uncertainties that can be introduced by machine saturation, temperature changes or loading variation.

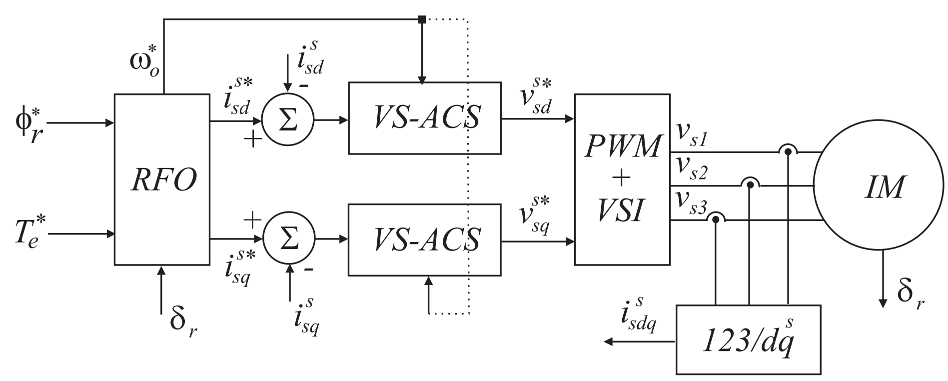


Fig. 2. Block diagram of the proposed *IM* motor drive system.

5.1 VS-MRAC Scheme

Consider that the linear first order plant of induction machine current-voltage transfer function W_{isdq}^s given by (67) and a reference model characterized by transfer function

$$M_{isdq}^s(s) = k_m \frac{N_m(s)}{D_m(s)} = \frac{b_e}{s + a_e}, \tag{68}$$

which attends for the stability constraints that is the constant b_s in (67) and b_e should have positive sign, as mentioned before. The output error can be defined as

$$e_{0sdq}^s = i_{sdq}^s - i_{mdq}^s, \tag{69}$$

where i^s_{mdq} (i^s_{md} and i^s_{mq}) are the outputs of the reference model. The tracking of the model control signal ($i^s_{sd} = i^s_{md}$ or $i^s_{sq} = i^s_{mq}$) is reached if the input of the control plant is defined as

$$v^s_{sdq} = \theta^*_{1dq} i^s_{sdq} + \theta^*_{2dq} i^{s*}_{sdq} \tag{70}$$

where θ^*_{1d} (θ^*_{1q}) and θ^*_{2d} (θ^*_{2q}) are the ideal controller parameters, that can be only determined if $W^s_{isdq}(s)$ is known. According to section 2, they can be determined as

$$\theta^*_{1d} = \theta^*_{1q} = \frac{a_s - a_e}{b_s}, \tag{71}$$

and

$$\theta^*_{2d} = \theta^*_{2q} = \frac{b_e}{b_s}. \tag{72}$$

Once $W^s_{isdq}(s)$ is not known, the controllers parameters $\theta_{1dq}(t)$ and $\theta_{2dq}(t)$ are updated by using switching laws as

$$\theta_{idq} = -\overline{\theta}_{idq} sgn(e^s_{0sdq} y^s_{isdq}) \tag{73}$$

where $i = [1,2]$ and y^s_{sdq} is the reference currents i^{s*}_{sdq} or the output currents i^s_{sdq}, and $\overline{\theta}_{idq} > \left|\theta^*_{idq}\right|$ are upper bounds which are assumed to be known, and the signal-function *sgn* is defined as

$$sgn(x) = \begin{cases} 1 & if \quad x > 0 \\ -1 & if \quad x < 0 \end{cases}. \tag{74}$$

Introducing nominal values of controller parameters $\theta_{idq(nom)}$ (ideally $\theta_{idq(nom)} = \theta^*_{idq}$). It is convenient to modify the control plant input given by (70) for the following

$$v_{sdq}^{s} = \theta^{T}\begin{bmatrix} v_{1dq} \\ i_{sdq}^{s} \\ v_{2dq} \\ i_{sdq}^{s*} \end{bmatrix} + \theta_{nom}^{T}\begin{bmatrix} i_{sdq}^{s} \\ i_{sdq}^{s*} \end{bmatrix}, \tag{75}$$

with $\theta^{T} = \begin{bmatrix} \theta_{v1dq} & \theta_{s1dq} & \theta_{v2dq} & \theta_{s2dq} \end{bmatrix}$, $\theta_{nom}^{T} = \begin{bmatrix} \theta_{s1dq(nom)} & \theta_{s2dq(nom)} \end{bmatrix}$ and

$$\begin{aligned} \dot{v}_{1dq} &= \Lambda v_{1dq} + v_{sdq}^{s} \\ \dot{v}_{2dq} &= \Lambda v_{2dq} + i_{sdq}^{s} \end{aligned}, \tag{76}$$

in which

$$\begin{aligned} \theta_{s1dq} &= -\overline{\theta}_{s1dq} sgn(e_{0sdq}^{s} i_{sdq}^{s}) + \theta_{s1dq(nom)} \\ \theta_{s2dq} &= -\overline{\theta}_{s2dq} sgn(e_{0sdq}^{s} i_{sdq}^{s*}) + \theta_{s2dq(nom)} \end{aligned}, \tag{77}$$

and

$$\begin{aligned} \theta_{v1dq} &= -\overline{\theta}_{v1dq} sgn(e_{0sdq}^{s} v_{1dq}) \\ \theta_{v2dq} &= -\overline{\theta}_{v2dq} sgn(e_{0sdq}^{s} v_{2dq}) \end{aligned}, \tag{78}$$

where θ_{s1dq}, θ_{s2dq}, θ_{v1dq} and θ_{v2dq} are the controller parameters, $\theta_{s1dq(nom)}$ and $\theta_{s2dq(nom)}$ are the nominal parameters of the controller, and v_{1dq} and v_{2dq} are the system plant input and output filtered signals, respectively. The constants $\overline{\theta}_{s1dq}$ or $\overline{\theta}_{s2dq}$ is chosen by considering that

$$\begin{aligned} \overline{\theta}_{s1dq} &> \left|\theta_{s1dq}^{*} - \theta_{s1dq(nom)}\right| \\ \overline{\theta}_{s2dq} &> \left|\theta_{s2dq}^{*} - \theta_{s2dq(nom)}\right| \end{aligned}, \tag{79}$$

The input and output filters given by equation (76) are designed as proposed in (Narendra & Annaswamy, 1989). The filter parameter Λ is chosen such that $N_{m}(s)$ is a factor of $\det(sI - \Lambda)$. Conventionally, these filters are used when the system plant is the second order or higher. However, it is used in the proposed controller to get two more parameters for minimizing the tracking error e_{0sdq}^{s}.

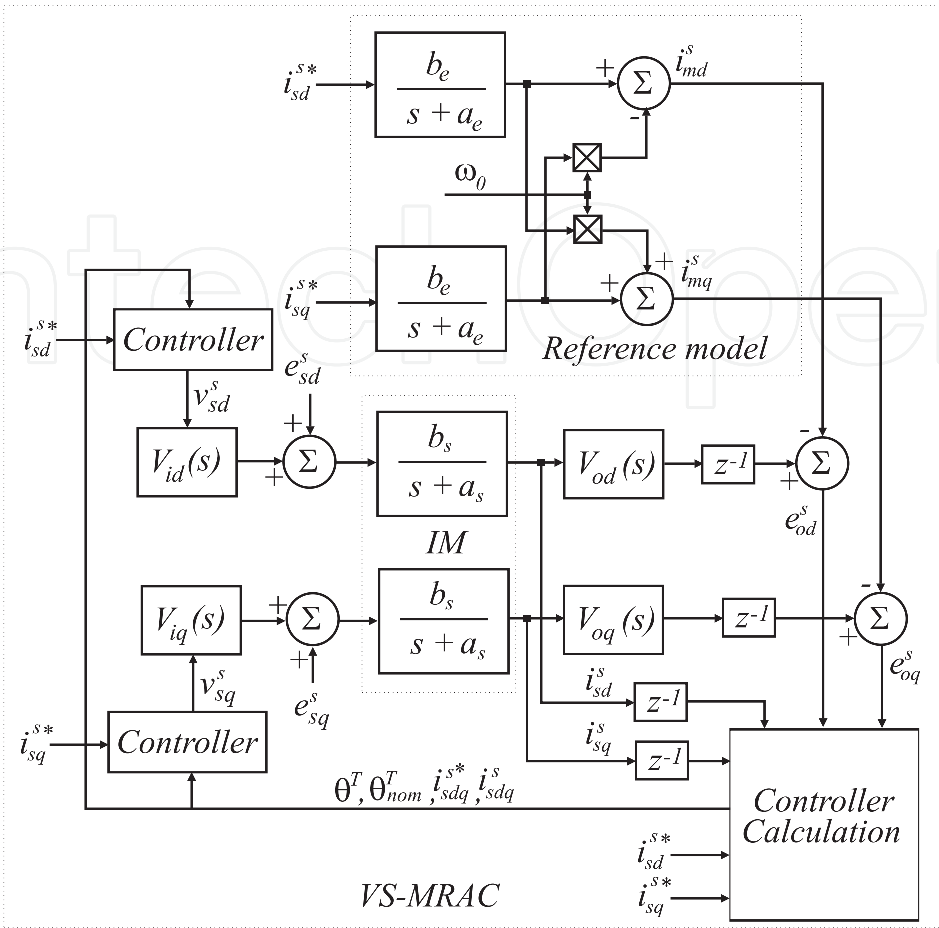


Fig. 3. Block diagram of proposed *VS-MRAC* current controller.

The block diagram of the *VS-MRAC* control algorithm is presented in Fig. 3. The proposed control scheme is composed by *VS* for calculating the controller parameters and a *MRAC* for determining the system desired performance. The *VS* is implemented by the block *Controller Calculation*, in which Equations (77) and (78) together are employed for determining θ_{s1dq}, θ_{s2dq}, θ_{v1dq} and θ_{v2dq}. These parameters are used by *Controller* blocks for generating the control signals v_{sdq}^{s}. To reduce the chattering at the output of controllers, input filters, represented by blocks $V_{id}(s)$ and $V_{iq}(s)$ are employed. They use filter model represented by Eqs. (76). These filtered voltages feed the *IM* which generates phase currents i_{sdq}^{s} which are also filtered by filter blocks $V_{od}(s)$ and $V_{oq}(s)$ and then, compared with the reference model output i_{mdq}^{s} for generating the output error e_{0sdq}^{s}. The reference models are implemented by two blocks which implements transfer functions (68). The output of these blocks is interconnected by coupling terms $-\omega_o I_{mq}^{s}$ and $\omega_o I_{md}^{s}$, respectively. This approach used to avoid the phase delay between the input (I_{sdq}^{s*}) and output (I_{mdq}^{s*}) of the reference model.

5.1.1 Design of the Controller

To design the proposed *VS-MRAC* controller, initially is necessary to choose a suitable reference model $M_{isdq}^{s}(s)$. Based on the parameters of the induction machine used in present study, given in Table 1, the reference model employed is

$$M_{isdq}^{s}(s) = \frac{550}{s + 550}, \tag{80}$$

From this reference model, the nominal values can be determined by using equations (71) and (72) which results in $\theta_{1sd(nom)} = \theta_{1sq(nom)} = 3.7$ and $\theta_{2sd(nom)} = \theta_{2sq(nom)} = 55$. Considering the restrictions given by (79), the parameters $\overline{\theta}_{s1dq}$ and $\overline{\theta}_{s2dq}$, chosen for achieving a control signal with minimum amplitude are $\overline{\theta}_{s1dq} = 0.37$ and $\overline{\theta}_{s2dq} = 5.5$. It is important to highlight that choice criteria determines how fast the system converges to their references. Moreover, it also determines the level of the chattering verified at the control system after its convergence. As mentioned before the use of input and output filters are not required for control plant of fist order. They are used here for smoothing the control signal. Their parameters was determined experimentally, which results in $\Lambda = 1, \theta_{v1d} = \theta_{v1d} = 2.0$ and $\theta_{v2d} = \theta_{v2q} = 0.1$. This solution is not unique and different adjust can be employed on these filters setup which addresses to different overall system performance.

5.2 VS-APPC Scheme

The first approach of *VS-APPC* in (Silva *et al.*, 2004) does not deal with unmodeled disturbances occurred at the system control loop like machine *fems*. To overcome this, a modified *VS-APPC* is proposed here.

Let us consider the first order *IM* current-voltage transfer function given by equation (67). The main objective is to estimate parameters a_s and b_s to generate the inputs v_{sd} and v_{sq} so that the machine phase currents i_{sd}^{s} and i_{sq}^{s} following their respective reference currents i_{sd}^{s*} and i_{sd}^{s*} and, the closed loop poles are assigned to those of a Hurwitz polynomials $A_s^*(s)$ given by

$$A^*(s) = s^3 + \alpha_2^* s^2 + \alpha_1^* s + \alpha_0^* , \tag{81}$$

where coefficients α_2^*, α_1^* and α_0^* determine the closed-loop performance requirements. To estimate the parameters a_s and b_s, the respective switching laws are used

$$\hat{a}_s = -\overline{a}_s sgn(e_{0sdq}^s i_{sdq}^s), \tag{82}$$

$$\hat{b}_s = \overline{b}_s sgn(e_{0sdq}^s v_{sdq}^s), \tag{83}$$

with the restrictions $\overline{a}_s > |a_s|$ and $\overline{b}_s > |b_s|$ satisfied, as mentioned before. The pole placements and the tracking objectives of proposed *VS-APPC* are achieved, if the following control law is employed

$$Q_m(s)L(s)V_{sdq}^s(s) = -P(s)(I_{sdq}^s - I_{sdq}^{s*}) \tag{84}$$

which addresses to the implementation of the controller transfer function

$$C_{sd}(s) = C_{sq}(s) = \frac{P(s)}{Q_m(s)L(s)}. \tag{85}$$

The polynomial $Q_m(s)$ is choose to satisfy $Q_m(s)I_{sd}^{s*}(s) = Q_m(s)I_{sq}^{s*}(s) = 0$. For the *IM* current-voltage control plant (see equation (67)) and considering that the *VS-APPC* control algorithms are implemented on the stator reference frame, which results in sinusoidal reference currents, a suitable choice for the controller polynomials are $Q_m(s) = s^2 + \omega_o^{*2}$ (internal model of sinusoidal reference signals i_{sd}^* and i_{sq}^*), $L(s) = 1$ and $P(s) = \hat{p}_2 s^2 + \hat{p}_1 s + \hat{p}_0$, where ω_o^* is the angular frequency of reference currents. This choice results in a current controller with the following transfer functions

$$C_{sd}(s) = C_{sq}(s) = \frac{\hat{p}_2 s^2 + \hat{p}_1 s + \hat{p}_0}{s^2 + \omega_o^{*2}} \tag{86}$$

where angular frequency ω_o^* is generated by vector *RFO* control scheme and coefficients $\hat{p}_2$, $\hat{p}_1$ and $\hat{p}_0$ are determined by solving the Diophantine equation for desired Hurwitz polynomial A_s^* (see equation (81)) as follows

$$\hat{p}_2 = \frac{\alpha_2^* - \hat{a}_s}{\hat{b}_s} \tag{87}$$

$$\hat{p}_1 = \frac{\alpha_1^* - \omega_o^{*2}}{\hat{b}_s} \tag{88}$$

$$\hat{p}_0 = \frac{\alpha_0^* - \omega_o^{*2}\hat{a}_s}{\hat{b}_s} \tag{89}$$

To avoid zero division on the equation (87)-(89), the switching law (83) is modified by

$$\hat{b}_s = \overline{b}_s sgn(e_{0sdq}^s v_{sdq}^s) + b_{s(nom)} \tag{90}$$

in which $b_{s(nom)}$ is the nominal values of b_s and the stability restriction becomes $\overline{b}_s > \left| b_s - b_{s(nom)} \right|$.

The control signals v_{sd}^s and v_{sq}^s generated at the output of the proposed controller *VS-APPC* can be derived from equation (86) which results in the following state-space model

$$\dot{x}_{1sdq}^s = x_{2sdq}^s + \hat{p}_1 \varepsilon_{sdq}^s \tag{91}$$

$$\dot{x}_{2sdq}^s = -\omega_o^2 x_{1sdq}^s + (\hat{p}_0 - \omega_o^2 \hat{p}_2)\varepsilon_{sdq}^s \tag{92}$$

$$v_{sdq}^s = x_{1sdq}^s + \hat{p}_2 \varepsilon_{sdq}^s \tag{93}$$

where $\varepsilon_{sdq}^s(t) = i_{sdq}^{s*} - i_{sdq}^s$ is the current error that is calculated from the measured quantities issued by data acquisition plug-in board as described next. Therefore, to generate the output signal of the controllers it is necessary to solve the equations (91)-(93).

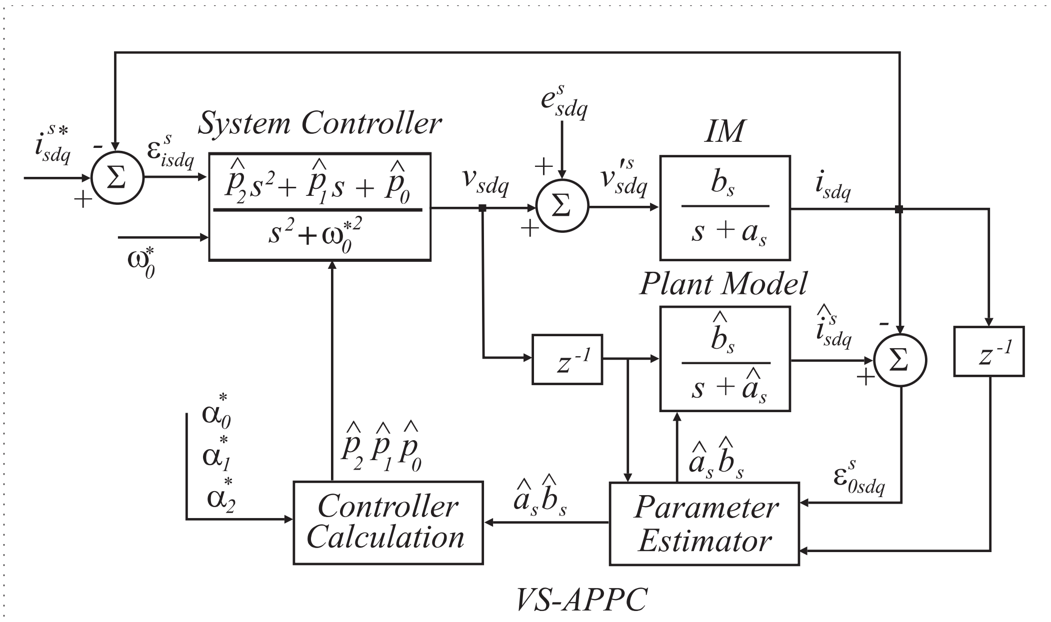


Fig. 4. Block diagram of proposed *VS-APPC* current controller.

The block diagram of the *VS-APPC* control algorithm for the machine current control loop is presented in Fig. 4. The proposed adaptive control scheme is composed a *SMC* parameter estimator and a machine current control loop subsystems. The *SMC* composed by blocks system controller and plant model identifies the dynamic of the *IM* current-voltage model. The output of this system generates the estimative of machine phase currents $\hat{i}^s_{sd}$ and $\hat{i}^s_{sq}$. The control loop subsystem composed by system controller and *IM* regulates the machine phase currents i^s_{sd} and i^s_{sq} and compensate the disturbances e^s_{sd} and e^s_{sq}. The comparison between the estimative currents ($\hat{i}^s_{sd}$ and $\hat{i}^s_{sq}$) and the machine phase currents (i^s_{sd} and i^s_{sq}) determines the estimation errors e^s_{0sd} and e^s_{0sq}. These errors together with machine voltages v^s_{sd} and v^s_{sq}, and *VS-APPC* algorithm set points $\overline{a}_s$, $\overline{b}_s$ and $\overline{b}_{s(nom)}$ are used for calculating parameter estimative $\hat{a}_s$ and $\hat{b}_s$, from the use of equations (82) and (90). These estimates update the plant model of the *IM* and are used by the controller calculation for together with, the coefficients of the desired polynomial A^*_s and angular frequency ω^*_o, determine the parameters of the system controller $\hat{p}_2$, $\hat{p}_1$ and $\hat{p}_0$. The introduction of the *IMP* into the controller modeling avoids the use of stator to synchronous reference frame transformations. With this approach, the robustness for unmodeled disturbances is achieved.

5.2.1 Design of the Controller

To design the proposed *VS-APPC* controller is necessary to choose a suitable polynomial and to determine the controllers coefficients $\hat{p}_2$, $\hat{p}_1$, and $\hat{p}_0$. A good choice criteria for accomplishing the bound system conditions, is to define a polynomial which roots are closed to the control plant time constants. The characteristics of *IM* used in this work are listed in the Table 1. The current-voltage transfer functions for *dq* phases are given by

$$\frac{I^s_{sdq}(s)}{V^s_{sdq}(s)} = \frac{10}{s+587} \tag{94}$$

A possible choice for suitable polynomial $A^*_s(s)$ can be

$$A^*_s(s) = (s+587)^3 \tag{95}$$

According to Equations (82), (90) and (87)-(89), and based on the desired polynomial (95), the estimative of the parameters of *VS-APPC* current controllers can be obtained as

$$\hat{p}_2 = \frac{1761 - \hat{a}_s}{\hat{b}_s} \tag{96}$$

$$\hat{p}_1 = \frac{1033707 - \omega_o^2}{\hat{b}_s} \tag{97}$$

$$\hat{p}_0 = \frac{202262003 - \omega_o^2 \hat{a}_s}{\hat{b}_s} \tag{98}$$

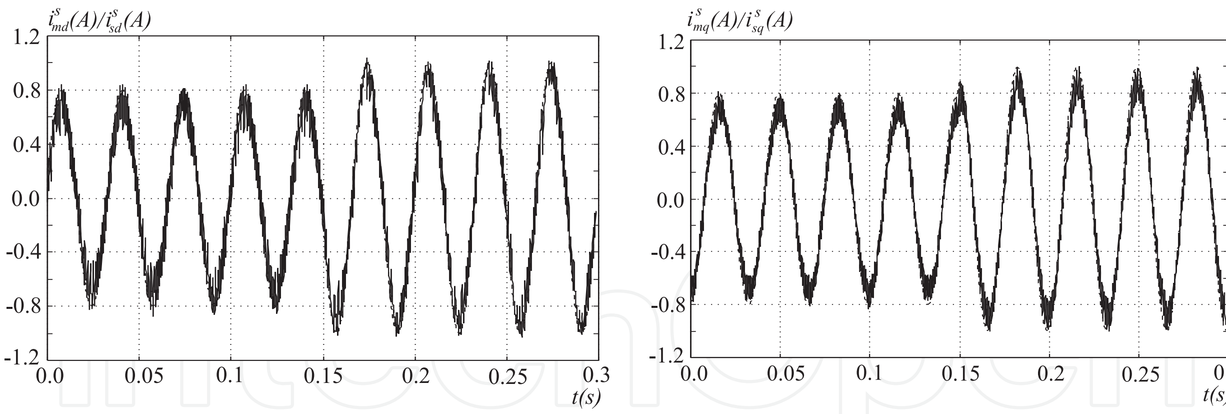
To define the coefficients of the switching laws it is necessary to take into account together the stability restrictions $\overline{a}_s > |a_s|$ and $\overline{b}_s > |b_s - b_{s(nom)}|$. Based on the simulation and the theoretical studies, it can be observed that the magnitude of the respective switching laws ($\overline{a}_s$ and $\overline{b}_s$) determine how fast the *VS-APPC* controllers converge to their respective references. However, the choice of greater values, results in controllers outputs (v_{sd} and v_{sq}) with high amplitudes, which can address to the operation of system with nonlinear behavior. Thus, a good design criteria is to choose the parameters closed to average values of control plant coefficients a_s and b_s. Using this design criteria for the *IM* employed in this work, the following values are obtained $b_{s(nom)} = 9$, $\overline{b}_s = 2$ and $\overline{a}_s = 600$. This solution is not unique and different design adjusts can be tested for different induction machines. The performance of these controllers is evaluated by simulation and experimental results as presented next.

$r_s = 31.0\Omega$	$r_r = 27.2\Omega$	$l_s = 0.8042H$	$l_r = 0.7992H$
$l_m = 0.7534H$	$J = 0.0133kg.m^2$	$F = 0.0146kg.m$	$P = 2$

Table 1. *IM* nominal parameters

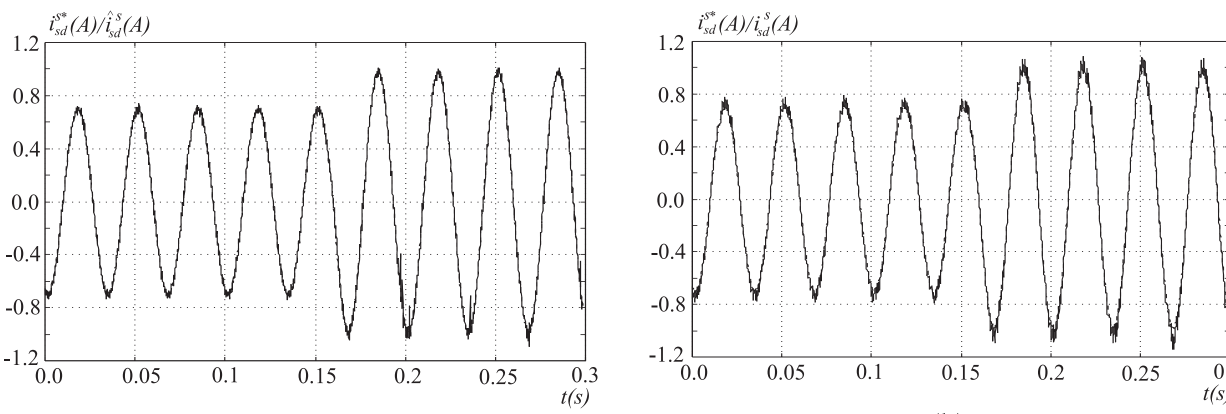
6. Experimental Results

The performance of the proposed *VS-MRAC* and *VS-APPC* adaptive controllers was evaluated by experimental results. To realize these tests, an experimental platform composed by a microcomputer equipped with a specific data acquisition card, a control board, *IM* and a three-phase power converter was used. The data of the *IM* used in this platform, are listed in Table 1. The command signals of three-phase power converter are generated by a microcomputer with a sampling time of 100 μs. The data acquisition card employs Hall effect sensors and *A/D* converters, connected to low-pass filters with cutoff frequency of $f_c = 2.5kHz$. Figures 5(a) and 5(b) show the experimental results of *VS-MRAC* control scheme. In these figures are present the graphs of the reference model phase currents i^s_{md} and i^s_{mq} superimposed to the machine phase currents i^s_{sd} and i^s_{sq}. In this experiment, the reference model currents are settled initially in $I^s_{mdq} = 0.8A$ and $f_s = 30Hz$. At the instant $t = 0.15s$, each reference model phase currents is changed by $I^s_{mdq} = 0.2A$. In these results it can be observed that the machine phase currents follow the model reference currents with a good transient response and a current ripple of $\Delta i^s_{sdq} \simeq 0.05A$. Figures 6-7 present the experimental results of *VS-APPC* control scheme. In the Fig. 6(a) are shown the graph of reference phase current i^{s*}_{sd} superimposed by its estimation phase current $\hat{i}^s_{sd}$. In this test, similar to the experiment realized to the *VS-MRAC*, the magnitude of the reference current is settled in $I^{s*}_{sdq} = 0.8A$ and at instant $t = 0.15s$, it is changed by $I^{s*}_{sdq} = 0.2A$. These results show that the estimation scheme employed in the *VS-APPC* estimates the machine phase current with small current ripple. Figure 6(b) shows the graphs of the reference phase current i^{s*}_{sd} superimposed by its corresponded machine phase current i^s_{sd}. In this result, it can be verified that the machine phase current converges to its reference current imposed by *RFO* vector control strategy. Similar to the results presented before, Fig. 7(a) presents the experimental results of reference phase current i^{s*}_{sq} superimposed by its estimation phase current $\hat{i}^s_{sq}$ and Fig. 7(b) shows the reference phase current i^{s*}_{sq} superimposed by its corresponded machine phase current i^s_{sq}. These results show that the *VS-APPC* also demonstrates a good performance. In comparison to the VS-MRAC, the machine phase currents of the *VS-APPC* present small current ripple.



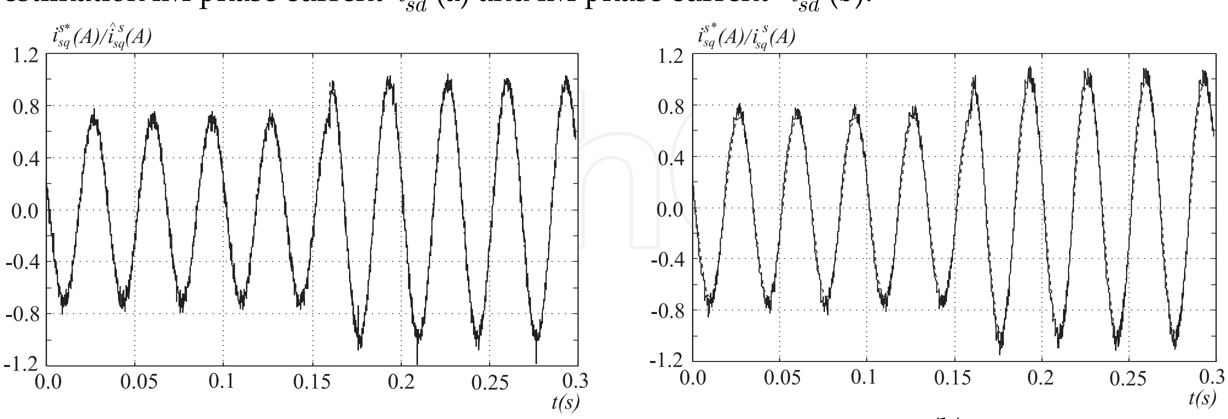
(a) (b)

Fig. 5. Experimental results of *VS-MRAC* phase currents i^s_{md} (a) and i^s_{mq} (b) superimposed to *IM* phase currents i^s_{sd} (a) and i^s_{sd} (b), respectively.



(a) (b)

Fig. 6. Experimental results of *VS-APPC* reference phase current i^{s*}_{sd} superimposed to estimation *IM* phase current $\hat{i}^s_{sd}$ (a) and *IM* phase current i^s_{sd} (b).



(a) (b)

Fig. 7. Experimental results of *VS-APPC* reference phase current i^{s*}_{sq} superimposed to estimation *IM* phase current $\hat{i}^s_{sq}$ (a) and *IM* phase current i^s_{sq} (b).

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Adaptive control has been a remarkable field for industrial and academic research since 1950s. Since more and more adaptive algorithms are applied in various control applications, it is becoming very important for practical implementation. As it can be confirmed from the increasing number of conferences and journals on adaptive control topics, it is certain that the adaptive control is a significant guidance for technology development.The authors the chapters in this book are professionals in their areas and their recent research results are presented in this book which will also provide new ideas for improved performance of various control application problems.

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